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Technical Report...

THIRD QUARTERLY REPORT

DEVELOPMENT OF C-BAND BEACON
FERRITE DUPLEXER
PHASE III

October, 1962

CONTRACT DA-36-039-SC-89158

Department of the Army Project
No. 3619-03-001

U. S. ARMY
SIGNAL RESEARCH AND DEVELOPMENT LABORATORY
FORT MONMOUTH, NEW JERSEY

SPERRY MICROWAVE ELECTRONICS COMPANY

CLEARWATER, FLORIDA

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DEVELOPMENT OF C-BAND BEACON
FERRITE DUPLEXER PHASE III

Third Quarterly Progress Report

31 July 1962 to 30 October 1962

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U. S. Army
Signal Research and Development Laboratory
Fort Monmouth, New Jersey

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DIVISION OF SPERRY RAND CORPORATION
CLEARWATER, FLORIDA

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1. PURPOSE AND SCOPE

The purpose and scope of this program was fully outlined in the first quarterly report.[1] The requirements set forth there remain unchanged.

2. ABSTRACT

The goal of this program is a C-band duplexer-limiter package meeting stringent electrical and mechanical requirements. The program has involved principally the investigation of four components: a four port coaxial junction circulator, a coaxial subsidiary resonance limiter, various forms of the multisphere gyromagnetic coupler, and a varactor limiter. Progress on all of these components is reported. A detailed theoretical analysis of the multisphere coupler and experimental results with the varactor limiter are presented in this report for the first time.

3. REFERENCES

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4. FACTUAL DATA

4.1. DESIGNS UNDER STUDY

The final form of the duplexer-limiter package to be developed on this program will be a well integrated combination of several components. The configurations in the block diagrams of Figure 1, of the second quarterly report, are still being studied. Work on the varactor limiter, which was purposely delayed during the early phases of the program, while awaiting the development of the subsidiary resonance limiter, is now well underway and valuable data are being obtained.

4.2 PROGRESS

4.2.1. Four-Port Circulator. During the last quarter various ferrite and garnet materials have been evaluated in a four port circulator. Emphasis has been placed on the use of garnet materials as they would more readily withstand the temperature and high power requirements of this program.

Figure 1 shows data obtained with a magnesium manganese ferrite material. Isolation of 15 to 20 db, VSWR of less than 1.2 and insertion loss of less than 1.5 db over a ten percent band have been obtained. This band is slightly higher in frequency than the design goal, but could be lowered with a modification of the puck diameter.

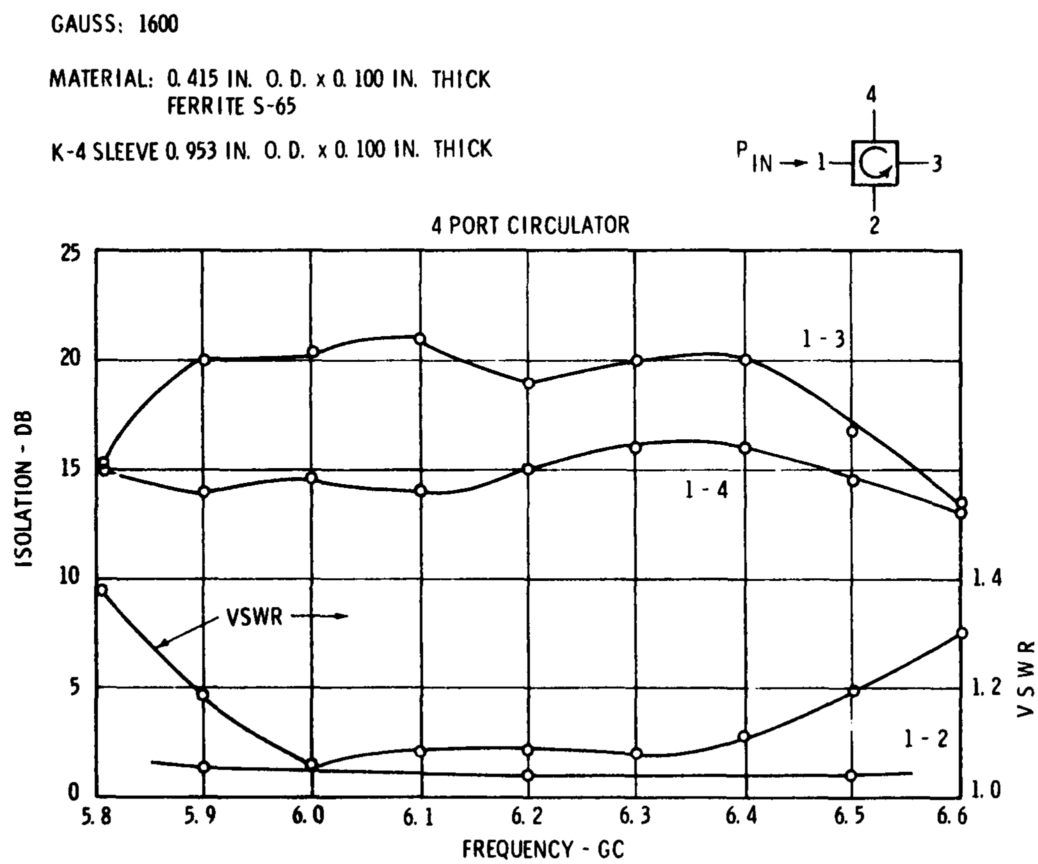


Figure 1. Electrical Characteristics of Four-Port Circulator (Magnesium Manganese Ferrite)

Figure 2 shows data obtained with a 60% gadilinium substituted yttrium iron garnet material having a linewidth of 250 oersteds. Isolation of 13 to 15 db, VSWR of less than 1.5 and insertion loss of less than 4 db over the band are obtained.

Figure 3 shows data obtained with yttrium iron garnet. Isolation of 19 to 35 db, VSWR of less than 1.7 and insertion loss of less than 5 db over the band have been obtained.

As can be seen from the curves, in going from a ferrite to a garnet, the insertion loss is increased. It can also be seen that the insertion loss increases more rapidly at the lower frequencies, indicating that the garnet is near resonance. Variation in VSWR and isolation between the ferrite and garnet loaded circulators is a result of the matching. Improved matching, however, would not greatly decrease the insertion loss.

At this point in the program the isolation and VSWR specification could be met with proper matching techniques but the insertion loss has presented a more formidable obstacle. In order to decrease the insertion loss, it will be necessary not only to find a material and configuration which will demonstrate intrinsically a lower loss, but also will have a circulation point farther from

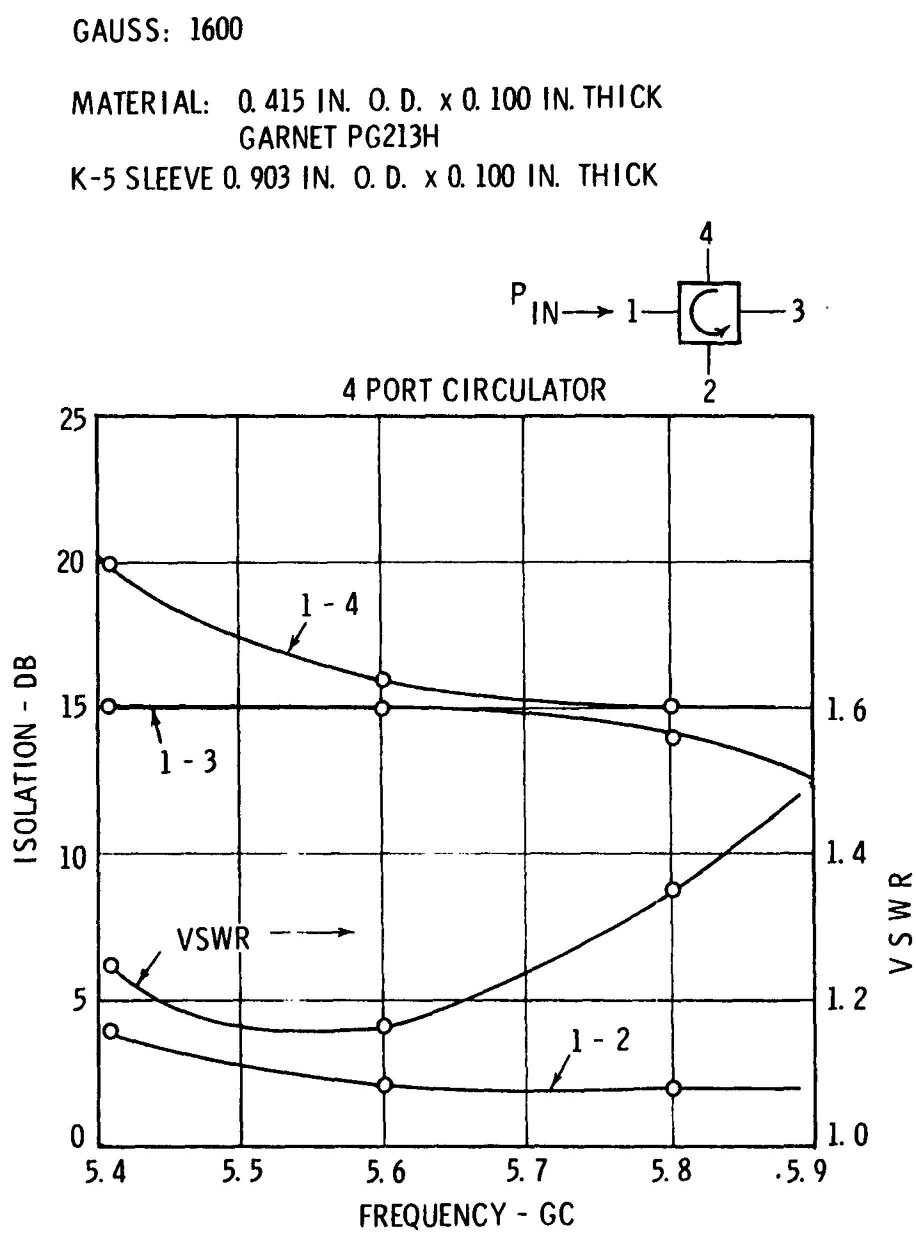


Figure 2. Electrical Characteristics of Four-Port Circulator (60% Gadilinium Substituted Garnet)

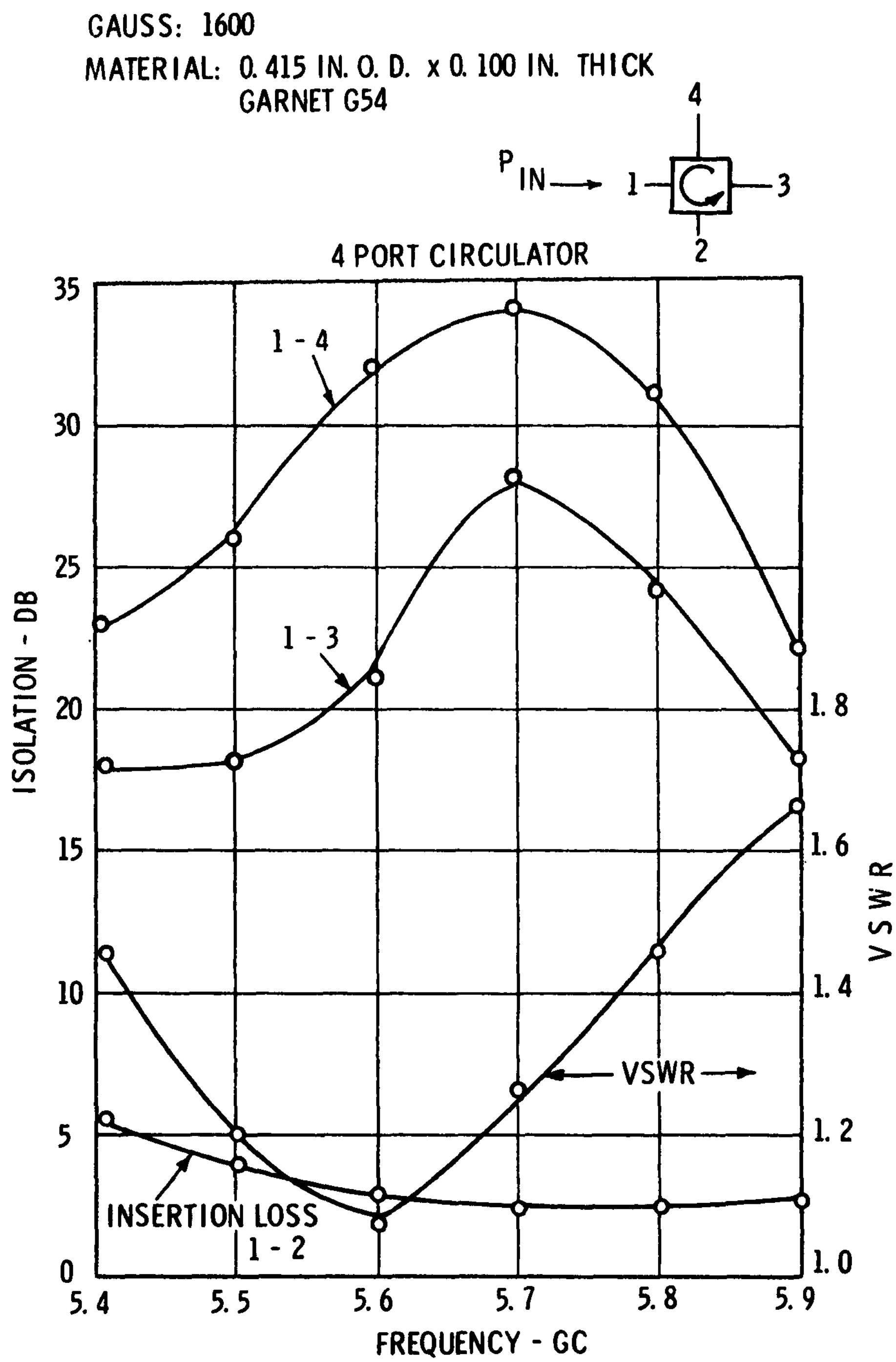


Figure 3. Electrical Characteristics of Four-Port Circulator (Yttrium Iron Garnet)

resonance.

The following four approaches were enacted: (1) A series of garnet pucks of varying diameters from 0.250 inches to 0.750 inches by 0.100 inches thick were fabricated. Variation in the garnet geometry changed the demagnetizing factors, allowing circulation to occur at a frequency removed from resonance. Evaluation of these materials showed that the insertion loss was still decreasing at a puck diameter of 0.750 inches. Larger garnet pucks are now being manufactured for evaluation. (2) Changes in the strip width and garnet puck height were made to see if a resonance peculiar to the geometry were present. No improvement was noted upon making the changes. (3) A housing utilizing a tuning screw on the axis of symmetry, as shown in Figure 4, was fabricated. Garnet pucks with a hole in the center to allow for the tuning screw adjustments were also fabricated. Comparison of this arrangement with a similar unit without the center hole and tuning screw showed no marked improvement. (4) Various garnet and ferrite materials of varying size were evaluated in an attempt to find a material with an overall lower insertion loss. This evaluation is still going on and as yet the minimum insertion loss encountered has

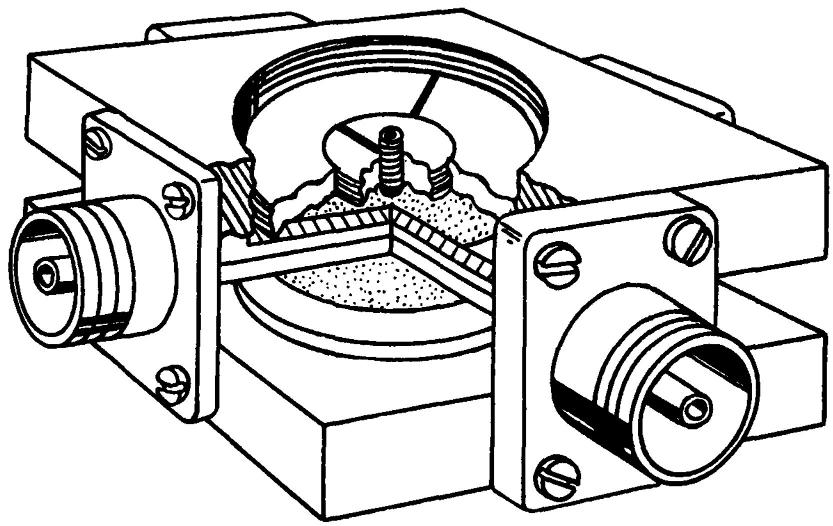


Figure 4. Four-Port Circulator Housing with Tuning Screw on Axis of Symmetry

been on the order of 1 db using a ferrite material.

During the next interval an evaluation of larger garnet pucks and continued evaluation of various materials is expected to lead to a decreased insertion loss. Greater emphasis will be placed on carefully matching the unit to see if asymmetries within the unit are contributing to the loss.

4.2.2 Multisphere Gyromagnetic Coupler Limiter

Early results obtained with multisphere gyromagnetic couplers have been somewhat disappointing. This has led, during the current reporting period, to a much more extensive theoretical analysis of the problems encountered in the design and construction of such a device.

A gyromagnetic coupling limiter (GCL) exhibits bandpass filter characteristics. The equivalent circuit of this component is closely approximated by a resonator coupled to a transmission line. The filter characteristics are determined by the degree of coupling and the magnitude of the unloaded Q of the resonator. The resonator in this particular case is a highly polished single crystal garnet sphere. The size of the sphere and its coupling parameters are chosen so as to yield the best compromise between loss and the desired limiting action above a certain signal power level. The limiter is designed primarily to act as a limiter and its bandpass characteristics which are usually of secondary interest are determined by the degree of coupling of the resonator to the transmission lines.

The application of conventional filter techniques involving resonators in series can be used in the analysis of broadband GCL. The major emphasis in filter design is usually placed on achieving a high signal rejection ratio below and above the cut off frequencies.

This aspect is of no direct concern in the design of a broadband limiter. A high signal rejection ratio at frequencies off the pass band will result as a natural consequence of the dimensioning of the composite filter. It should be mentioned here that a single resonator limiter covering a pass band of 500 Mc at a constant insertion loss of approximately 0.5 db would satisfy the pass band requirements of the component to be developed. A broadband single resonator gyromagnetic limiter can be designed by selecting a broad linewidth resonator. The insertion loss, however, would be excessive as the dissipative loss increases with the increasing linewidth of the ferrimagnetic solid.

A combination of a multiple of narrow linewidth resonators is obviously the next consideration. A Chebishev rather than a Butterworth filter design has been investigated.

A filter consisting of an even number of resonators would require a narrower bandwidth for the individual resonators than would be the case where a single resonator is designed to cover the entire pass band.

Analyzing the filter concept further, assuming a pass band of constant insertion loss reaching from 5.4 to 5.9 KMC, a 3 db bandwidth of 600 Mc can be extrapolated. From a further extrapolation, the 25 db bandwidth is found to be in the order of 800 Mc. It should be pointed out that the analysis does not take into account insertion loss due to any absorptive mechanism, such as loss in the garnet material.

Allowing a ripple of 0.1 db in the band pass, the number of necessary resonators can be found analytically or simpler by means of calculated tables[24].

A "6-pole" network of a constant K configuration will yield the desired band pass characteristics. For the case of a band pass filter, the number of poles describes the number of necessary resonators. The individual resonators are tuned to the center frequency of the pass band (geometric means of the upper and lower pass band cut-off frequencies).

The internal coupling of each individual resonator is optimized and the Q of the resonators is controlled by the proper selection of the linewidth of the ferrimagnetic resonator spheres. This step avoids the necessity of additional coupling networks between each resonator. Further analysis shows that the minimum Q requirements can be easily satisfied with gyromagnetic resonators. The bandwidth requirements imposed on the individual resonators, however, are still prohibitive and can only be obtained with a relatively broad linewidth material. A broad line-width in turn will always result in a high loss that will not appear in the calculations of the over-all frequency response.

The matter of loss is actually the crucial problem with this approach. While the filter approach will satisfy the response requirements, the loss will be prohibitively high. An estimate, based on experimental results obtained over the past several years at Sperry Microwave Electronics Company, places this loss at about 3 db. A loss of this magnitude is considered prohibitive for the purposes of this program.

An alternate approach involves the use of a parallel arrangement of resonators. This is actually the approach that has received the most experimental attention. The serpentine line, the multisphere common ground plane configuration, and the stagger tuned multistage GCL are all parallel connected.

Thus far in the program, relatively poor experimental results have been obtained with these devices. All three forms have been investigated in the course of the program. During the present reporting period a more thorough theoretical investigation of the problem of the multisphere coupler was initiated; and, as the investigation progressed, the common ground plane form of the device was selected for experimentation. This structure is felt to be the one that is most closely approximated by the theory; and has been found to have an advantage in mechanical simplicity over the other forms of the device, which is a considerable help in experimentation. A photograph of the test piece presently being used is presented in Figure 5.

The theoretical investigation is presented in Appendix A. It is essentially an equivalent circuit development of a 2-sphere model of the multisphere coupler. The final result is quite complex, and the use of a computer would apparently be necessary for a thorough analysis. Whether or not this result is of sufficient value to warrant the required computer work is a matter that will be decided early in the next quarter.

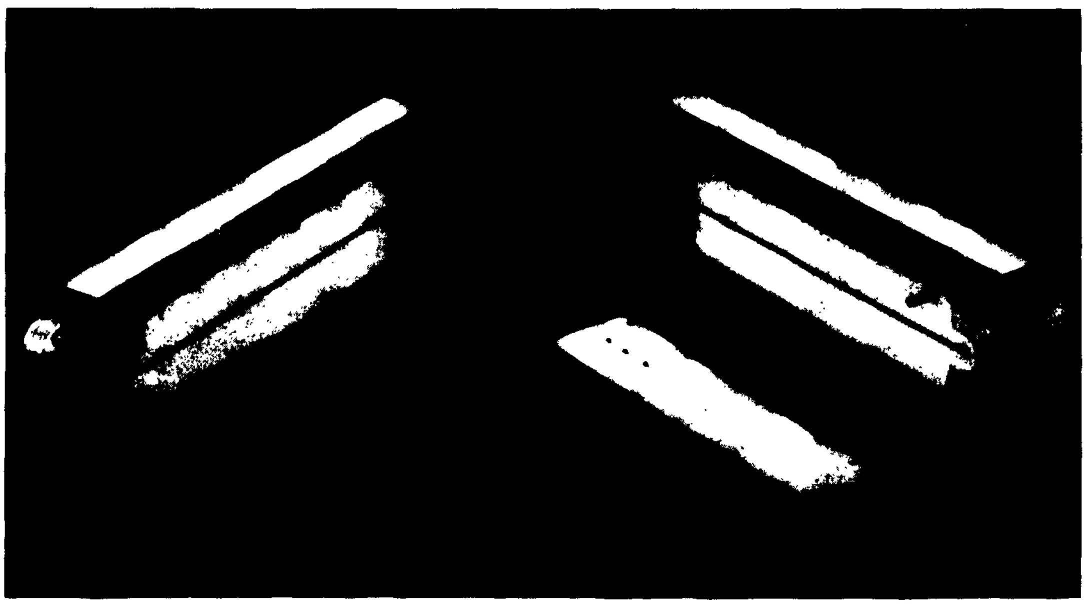


Figure 5. Recent Model of the Common Ground Plane Multisphere Coupler

Hopefully, results can be derived that will lead to more fruitful work in the laboratory.

4.2.3 Subsidiary Resonance Limiter. Material investigations have comprised a large part of the subsidiary resonance limiter investigation up to this time. YIG poly crystaline material has received the most attention. It has been shown that both narrow line width and very low loss tangent are required for good limiting and acceptable insertion loss. The application of improved materials preparation techniques at Sperry has recently resulted in the production of a garnet material, having a linewidth of less than 30 oe and a loss tanget of less than 0.0001. This material appears to be usable in the subsidiary resonance limiter. It is now available in small quantities for development work.

With the material problem close to solution, considerable effort has been expended during this quarter in developing an optimum limiter configuration in strip line. The goal of this work is a device with a flat leakage in the neighborhood of 20 watts and a low power loss of less than 0.5 db. This is felt to be an ambitious goal, and present performance, while good, does not yet match it. A limiting curve representative of the performance obtained to date is illustrated in Figure 6. The

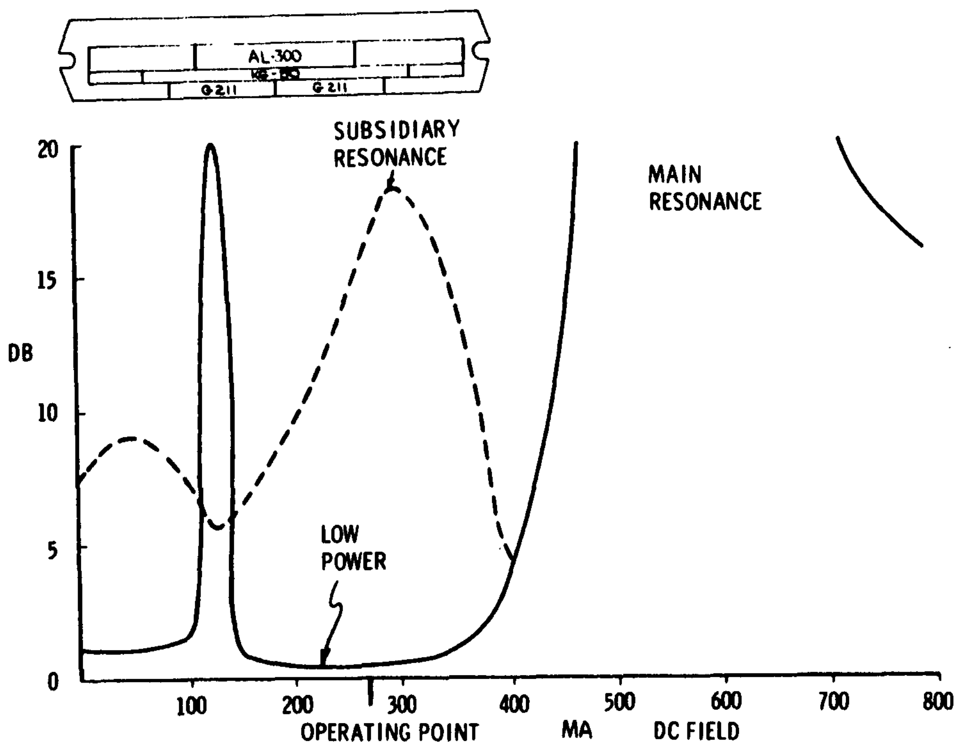


Figure 6. Electrical Characteristics of the Coaxial Subsidiary Resonance Limiter at 5.6 Gc

maximum loss of the device over the 5.4 to 5.9 kmc range is 0.8 db. The material configuration employed is illustrated in the Figure. Low loss, a maximum of less than 0.5 has been obtained in a slightly different configuration; but this was obtained at the expense of approximately 4 db of limiting.

Experimental results indicate that both lower loss and lower threshold are obtainable in a device of this type over a narrow band. Work in the immediate future will be concerned with the development of a configuration which leads to broad band, high limiting and low loss.

4.2.4 <u>Varactor Limiter</u>. "Clean up" varactor limiters have received considerable attention during this reporting period. The basic circuit under investigation is pictured in Figure B-1 of Appendix B. The design of these devices is more straight forward and amenable to calculation than the previously discussed ferrite components. A theoretical analysis of a simple varactor limiter is presented in an Appendix. These calculations lead to valuable design information.

The physical structure being used in pictured in Figure 7. The indicated short is reflected as a high impedence in parallel with the varactor. Pill varactors are being used to provide minimum lead inductance. RCA

Figure 7. The Varactor Limiter

varactors with the following characteristics have proven most successful:

$$f_c \approx 220 \text{ gc } (-6V)$$

$$c_j \approx 0.5 \text{ pf } (-6V)$$

$$R_s \approx 1.5 \text{ ohms}$$

The limiter structure was specifically designed to allow two such units to be placed back to back with a $\lambda/4$ spacing between the stubs at the mid band of the 5.4 to 5.9 gc band. It can be shown that a $\lambda/4$ spacing will yield a dynamic range that is approximately the sum of the dynamic ranges of the two limiting stages plus an additional factor of perhaps 6 db.

The structure was also designed to permit the use of two varactors in a single stage of limiting. This provision was made with the knowledge that the leakage energy from the subsidiary resonance limiter may be excessively high for a single varactor. It has been found that the best results can be obtained when both varactors are placed on the same side of the strip.

Essentially the same values of high power isolation are obtained with both the single varactor and the two varactor versions of the device but the loss is roughly doubled in the two varactor version. For a maximum high power isolation of 8 db the loss in the single varactor limiter is about 0.2 db. As indicated in appendix, increased values of high power isolation are obtained at the expense of higher low power insertion loss. A two varactor limiter with a high power isolation of 15 db exhibits loss in the neighborhood of 1 db. This configuration has been tested to 1 Kw with a 0.001 duty cycle, behind a subsidiary resonance limiter with a maximum high power attenuation of 14 db. No deterioration in performance of the varactor limiter was detected, and it is assumed that the varactors suffered no high power damage. Higher power tests of this type leading eventually to a test with 4 KW into the subsidiary resonance limiter are planned for the near future.

Assuming 4 kw into the limiter as the extreme high power condition and assuming essentially no spike attenuation in the subsidiary resonance limiter, the varactor limiter will have to handle a 4 Kw spike 20 to 30 nanoseconds in half width. An acceptable performance would be a reduction in spike height to a level of about 1 watt. This requires at least 36 db of high power attenuation in the varactor limiter alone. On the basis of present operation this level of high power attenuation will be accompanied by excessive low power loss, unless the balance between high power and low power loss in the present device is improved. This improvement is the goal of work in the immediate future.

5. CONCLUSIONS

The lack of promising results with any of the forms of the multisphere gyromagnetic coupler has forced the adoption of the subsidiary resonance limiter approach as the principle line of limiter investigation for the program. Information forthcoming from a more thorough theoretical investigation of the multisphere coupler could possibly reverse this decision.

High insertion loss in all of the components involved now appears to be the largest obstacle to a successful conclusion of the program. The last quarter should see considerable improvement in this characteristic, but the possibility of meeting the program's design goal in antenna to receiver insertion loss now appears small.

6. PROGRAM FOR THE NEXT QUARTER

Continued attention will be given to the problem of insertion loss in the 4-port circulator, and the behavior of the device at high power will be studied.

Theoretical investigations and some experimental work on the multisphere gyromagnetic coupler will continue.

The 4-port circulator, the coaxial subsidiary resonance limiter, and the varactor limiter will be combined. This should take place during the second month of the quarter, and will probably be the configuration to be delivered as the program's final device. Emphasis will be on high power performance and on the task of obtaining lower antenna to receiver insertion loss.

7. DISTRIBUTION OF MANPOWER HOURS

The following is the distribution of manpower hours for the present reporting period:

B. J. Duncan	50
J. Clark	256
E. W. Matthews	9
J. Brown	67
G. Neumann	20
D. H. Landry	163
R. E. Willoughby	6
J. C. Hoover	8
G. V. Buehler	175

The manpower hours for July which were not given in the 2nd quarterly report because of administrative difficulties are presented below:

B. J. Duncan	5
J. Clark	127
E. W. Matthews	6
J. Brown	33
G. Neumann	4
D. H. Landry	52
R. E. Willoughby	57
J. C. Hoover	23
W. C. Heithaus	1

8. IDENTIFICATION OF KEY TECHNICAL PERSONNEL

The key technical personnel assigned to the program have remained unchanged since the submission of the last quarterly report.

APPENDIX A

PARALLEL CONNECTION OF GYROMAGNETIC COUPLING LIMITERS

INTRODUCTION

A gyromagnetic coupler limiter utilizing two gyromagnetic resonators is analyzed from an equivalent circuit viewpoint. Below the threshold, the limiter can be regarded as a symmetrical and reciprocal network. The latter statement is not entirely true since the signal is subjected to a different phase shift depending on the direction of the signal with respect to the direction of the applied magnetic field. Measurements, however, have shown that the differential phase shift of the signal is constant (180°) as a function of frequency. Therefore the theory of reciprocal symmetrical networks will apply.

A number of gyromagnetic couplers might be connected in parallel and spaced such as to yield a broader bandwidth than that of a single resonator. This parallel connection has limitations because it requires the use of frequency sensitive impedance transformers.

Figure A-1 shows the circuit concept for three resonators, each tuned to a different resonant frequency. The spacing (d) between the transmission line I and II may be neglected as the device can be designed such that the spacing is reduced to about twice the sphere (resonator) diameter. The spacing between the resonators is critical

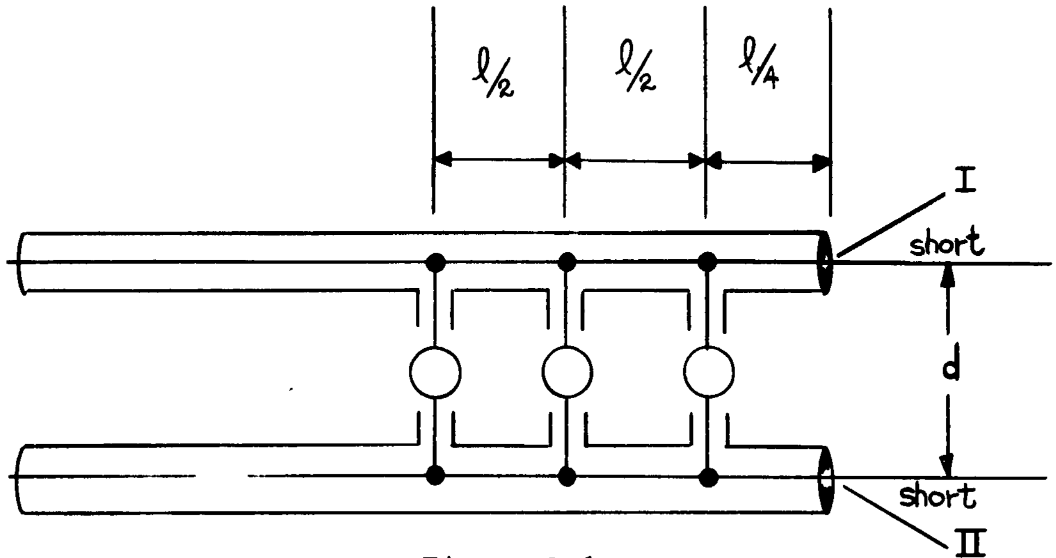
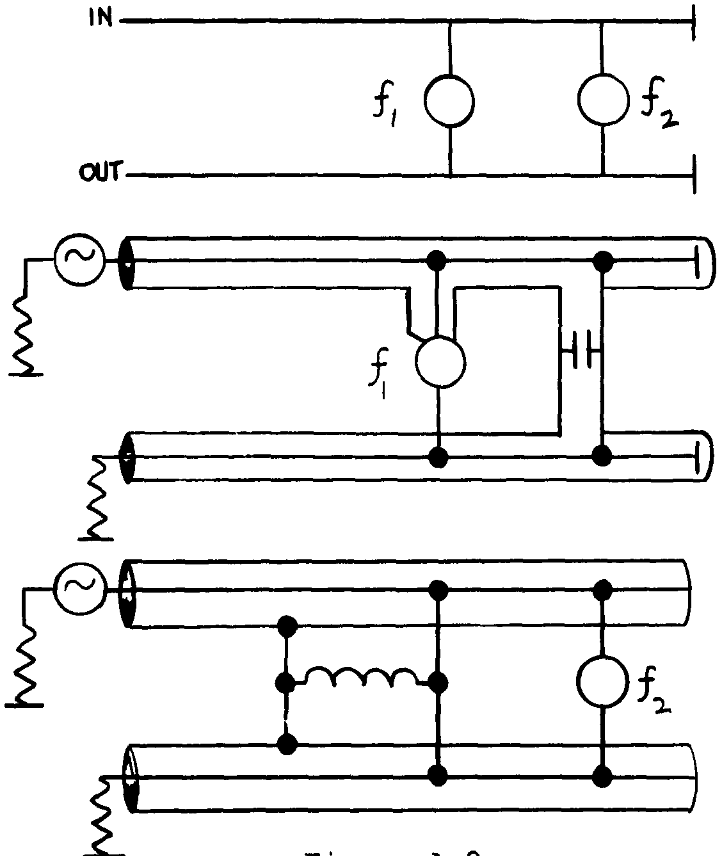


Figure A-1

for a minimum reflective loss within the passband of the device. Assuming the center frequency of the passband to be f_o, the resonant frequencies are chosen symmetrically above and below the center frequency of the passband.

The resonator tuned for a frequency higher than the band center frequency will present a shunt susceptance at frequencies below. If the signal frequency is higher than the band center frequency, the resonators with lower resonant frequencies will present a shunt susceptance of opposite sign. Figure A-2 illustrates this concept. These shunt susceptances alter the electrical lengths of the lines with which the resonators are spaced. The electrical length of the terminating



f_o = band $\quad \sqrt{f_1 \, f_2} = f_o$

$f_1 < f_o$

$f_2 > f_o$

Note: The resonators are treated as 4-terminal networks of zero electrical length. The necessary impedance transformers for creating a current maximum at the resonator are included in the 4-terminal network

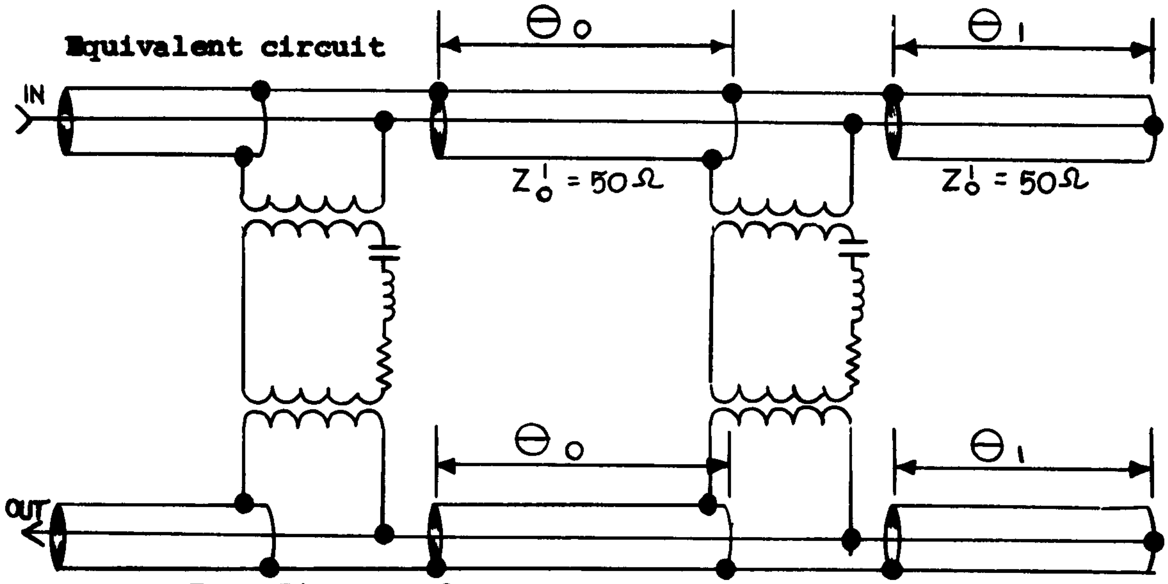
Figure A-2

$\lambda/4$ transformer is also a function of frequency. It would be desirable to know over what frequency band a parallel configuration will operate without exceeding a given amount of reflection loss and what the spacings between the resonators should be and how long the terminating transformer must be to obtain a maximum band pass at a tolerable VSWR.

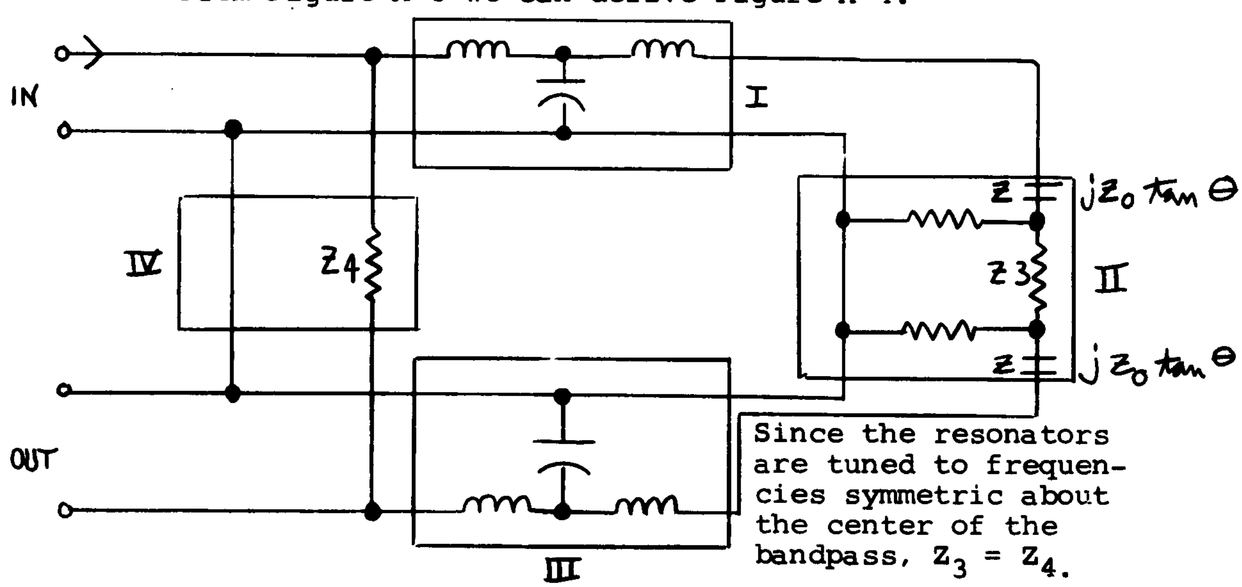
A structure incorporating 2 resonators with resonant frequencies f_1 and f_2 spaced symmetrically with respect to the center frequency (f) of the pandpass is analyzed. Under this condition the magnitude of the impedance (or susceptance) of individual resonators at frequencies $+\Delta f$ and $-\Delta f$ removed from f is equal and differs in sign only.

ANALYSIS OF A TWO RESONATOR STRUCTURE

Equivalent circuit



From Figure A-3 we can derive Figure A-4.



Since the resonators are tuned to frequencies symmetric about the center of the bandpass, $Z_3 = Z_4$.

Figure 3. Top
Figure 4. Bottom

The resonators are assumed to be lossless. The impedances of the stubs can be considered with the resonator impedance Z_3 to form a 4-terminal network II, so that the equivalent circuit of the two-resonator limiter can be split up into four 4-terminal networks as illustrated in Figure A-4.

Network I and III have a transmission matrix of the form

$$\left| a_I \right| = \left| a_{III} \right| = \begin{vmatrix} \cos\theta_o & j\, Z_o' \sin\theta_o \\ j\, \dfrac{\sin\theta_o}{Z_o'} & \cos\theta_o \end{vmatrix}$$

where Z_o' = characteristic impedance of the transmission line. For both networks the length of the transformer is chosen to be equal (practical design aspect).

The transmission matrix of network II can be prescribed as follows.

$$\left| a_{II} \right| = \begin{vmatrix} \dfrac{Z + Z_3}{Z} & Z_3 \\ \dfrac{2Z + Z_3}{Z^2} & \dfrac{Z + Z_3}{Z} \end{vmatrix}$$

The four-terminal networks I, II, III can now be merged into one by multiplication of the three transmission matrices. The transmission matrix of this new 4-terminal network is:

$$\left| A_{I\ II\ III} \right| = \begin{vmatrix} M_{11} & M_{12} \\ M_{21} & M_{22} \end{vmatrix}$$

where

$$M_{11} = (\cos\theta_o A_{11} + jZ_o' \sin\theta_o A_{21})\cos\theta_o + (\cos\theta_o A_{12} + jZ_o' \sin\theta_o A_{22})\, j\frac{\sin\theta_o}{Z_o'}$$

$$M_{12} = (\cos\theta_o A_{11} + jZ_o' \sin\theta_o A_{21})jZ_o' \sin\theta_o + (\cos\theta_o A_{12} + jZ_o' \sin\theta_o A_{22})\cos\theta_o$$

$$M_{21} = (j\frac{\sin\theta_o}{Z_o'} A_{11} + \cos\theta_o A_{21})\cos\theta_o + (j\frac{\sin\theta_o A_{12}}{Z_o'} + \cos\theta_o A_{22})j\frac{\sin\theta_o}{Z_o'}$$

$$M_{22} = (j\frac{\sin\theta_o}{Z_o'} A_{11} + \cos\theta_o A_{21})j\sin\theta_o Z_o' + (j\frac{\sin\theta_o}{Z_o'} A_{12} + \cos\theta_o A_{22})\cos\theta_o$$

$$A_{11} = \frac{Z + Z_3}{Z}$$

$$A_{12} = Z_3$$

$$A_{21} = \frac{2Z + Z_3}{Z^2}$$

$$A_{22} = \frac{Z + Z_3}{Z} = A$$

θ_o = electrical length of the transmission line transformer separating the two resonators

Z = impedance of the transmission line transformer

Z_3 and Z_4 = resonator impedance.

The admittance matrix for the combined network (I, II, III, in tandem) can be found straight-forwardly from the transmission matrix. By adding this admittance matrix to the admittance matrix for network IV and then transforming back to a transmission matrix, the square of the characteristic impedance of network IV in parallel with network I, II, and III in cascade is found to be:

$$Z_o^2 = \frac{M_{12}^2 Z_4^2}{Z_4^2 + 2Z_4 - Z_4^2 M_{22}^2 - 2M_{12}M_{22}Z_4}$$

From this expression the VSWR of the network can be calculated for any frequency since the VSWR can be expressed as

$$\frac{Z_o'}{Z_o} \text{ or } \frac{Z_o}{Z_o'} .$$

Substituting the expressions for M_{12} and M_{22} in the above equation and rationalizing all reactive impedances, a real expression for Z_o^2 is found. This expression is the exact solution for Z_o^2. For convenience the rationalized numerator and denominator of this lengthy expression are presented separately.

Rationalized Numerator

$$Z_3^2 Z_4^2 \cos^4\theta_o + \sin^2\theta_o \cos^2\theta_o (4Z_o^2 A^2 Z_4^2 - 2Z_3 Z_4^2 Z_o^{2\prime} A_{21})$$
$$+ Z_o^{4\prime} Z_4^2 A_{21}^2 \sin^4\theta + 4AZ_o' Z_4^2 \cos\theta \sin\theta (Z_3 \cos^2\theta - Z_o^2 A_{21} \sin^2\theta)$$

Rationalized Denominator

$-Z_4{}^2$

$-\cos^2\theta(+2Z_3Z_4)$

$+\sin^2\theta(-2Z_o^{2\prime}Z_4A_{21})$

$+\sin^4\theta(-2A_{21}Z_o{}^2Z_4A + Z_4{}^2A^2)$

$+\cos^4\theta(Z_4{}^2A^2 + 2AZ_3Z_4)$

$-\sin\theta\cos\theta(4Z_4Z_o{}'A)$

$$+\sin^2\theta\cos^2\theta\Big(-2Z_4Z_3A + Z_4{}^2Z_oA_{21}{}^2 + \frac{Z_4{}^2Z_3{}^2}{Z_o{}^2} - 2A_{21}Z_3Z_4{}^2$$

$$- 2Z_4{}^2A^2 + 2Z_o^{2\prime}A_{21}AZ_4 + 4A_{21}Z_o^{2\prime}Z_4A - 4Z_3Z_4A\Big)$$

$$+\sin^3\theta\cos\theta\Big(-2A_{21}{}^2Z_o{}'^3Z_4 - 4A^2Z_4Z_o{}' + 2Z_4{}^2A\ \frac{Z_3}{Z_o} - 2Z_4{}^2AA_{21}Z_o{}'\Big)$$

$$+\cos^3\theta\sin\theta\Big(2Z_4{}^2AA_{21}Z_o{}' - 2Z_4{}^2A\ \frac{Z_3}{Z_o} + 4A^2Z_4Z_o{}' + 2A_{21}Z_o{}'Z_3Z_4$$

$$+ 2\ \frac{Z_3{}^2}{Z_o}\ Z_4\Big)$$

$Z_o{}'$ = characteristic impedence of the transmission line transformer

Z_3, Z_4 = magnitude of resonator impedence

$A = A_{22} = A_{11} = \dfrac{Z + Z_3}{Z}$ where $Z = Z_o{}'\tan\theta$

$\theta = \dfrac{2\pi \cdot l}{\lambda}$

$\theta_o = \dfrac{2\pi \cdot l_o}{\lambda}$

Resonator Impedance

For the evaluation of the general expression of Z_o^2 the resonator impedances Z_4 and Z_3 have to be defined as a function of frequency. This can be accomplished by analyzing the passband characteristics of a gyromagnetic coupling resonator.

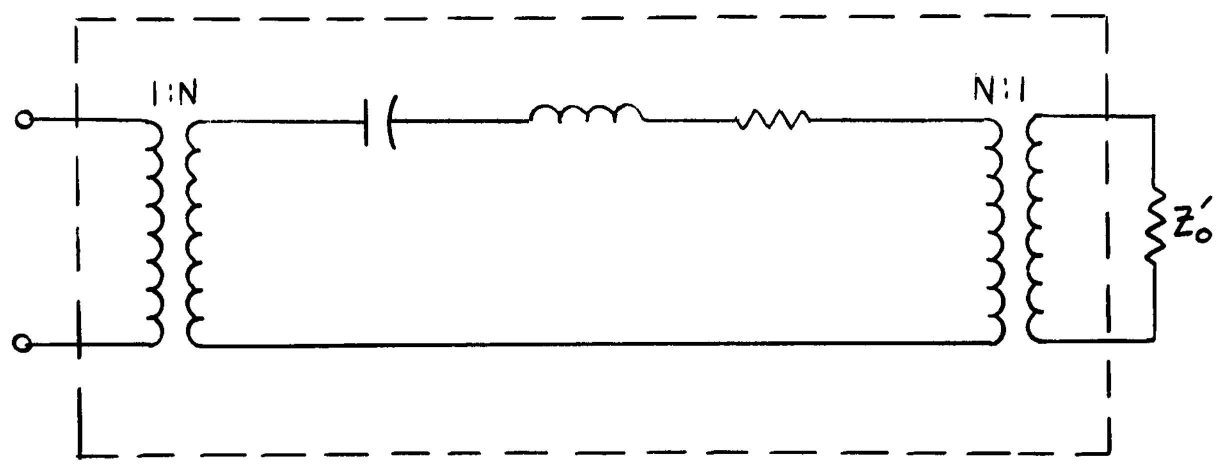


Figure A-5

The following expression for the attenuation through a single resonator gyromagnetic coupler limiter for power levels below the threshold was derived in Reference 3 from the equivalent circuit shown in Figure A-5.

$$\alpha_{db} = 10 \log_{10} \left[1 + Q_L^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2 \right]$$

where Q_L is the loaded Q of the resonator.

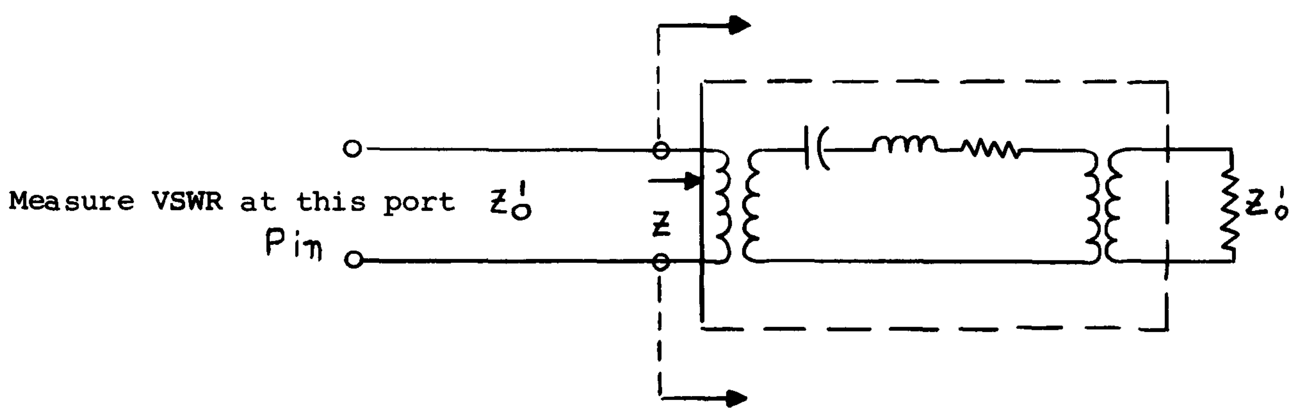
The resonator is assumed to have a very high unloaded Q so that the insertion loss α is due primarily to reflection loss. Reflection loss can be written in terms of VSWR as

$$\text{Reflection loss} = \frac{(S + 1)^2}{4S}$$

$$(\text{reflection loss})_{db} = 10 \log_{10} \frac{(S + 1)^2}{4S}$$

$$\alpha_{db} = (\text{reflection loss})_{db}$$

$$\therefore \; 1 + Q_L{}^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2 = \frac{(S + 1)^2}{4S}$$



$$S = \frac{Z_o'}{Z} \text{ or } \frac{Z}{Z_o'} = \text{VSWR}$$

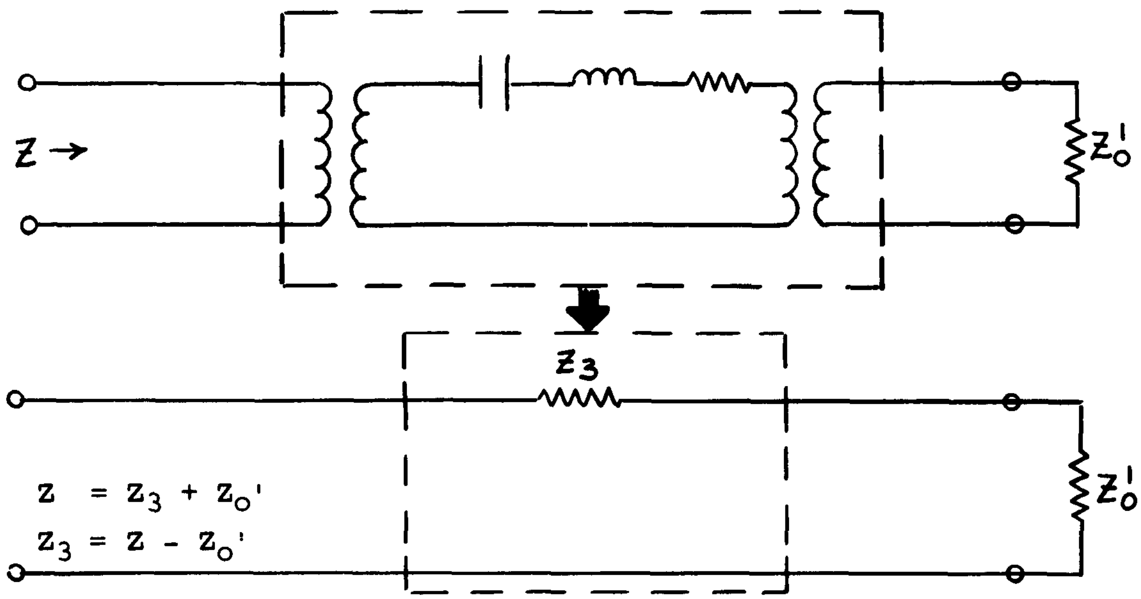
$$1 + Q_L{}^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2 \qquad \frac{\left(\frac{Z_o'}{Z} + 1 \right)^2}{4 \, \frac{Z_o'}{Z}} \qquad \frac{(Z_o' + Z)^2}{4ZZ_o'}$$

Let $Q_L^{\,2}\left(\frac{f}{f_o} - \frac{f_o}{f}\right)^2 = P$

then

$$Z = Z_o'\left[1 + 2P \pm 2\sqrt{P(1 + P)}\right] .$$

This Z is the impedance looking into the resonator which is terminated in Z_o'. To find a series equivalent impedance for the resonator, consider the following 4-terminal networks.



$$Z = Z_3 + Z_o'$$

$$Z_3 = Z - Z_o'$$

$$Z_3 = Z_o'\left(1 + 2P \pm 2\sqrt{P(1 + P)}\right) - Z_o'$$

$$Z_3 = Z_o'\left(2P \pm 2\sqrt{P(1 + P)}\right)$$ = series equivalent impedance of resonator

$$Z_3 = 2Z_o'\left(P \pm \sqrt{P(1 + P)}\right).$$

Approximations at the half power points indicate that the minus sign should be eleminated. Substituting for P,

$$Z_3 = 2Z_o' \left[Q_L^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2 + \sqrt{Q_L^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2 \left\{ 1 + Q_L^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2 \right\}} \right]$$

$$Z_3 = 2Z_o' \left[Q_L \left(\frac{f}{f_o} - \frac{f_o}{f} \right) \right] \left[Q_L \left(\frac{f}{f_o} - \frac{f_o}{f} \right) + \sqrt{1 + Q_L^2 \left(\frac{f}{f_o} - \frac{f_o}{f} \right)^2} \right] .$$

APPENDIX B

VARACTOR ON A QUARTER WAVE STUB

INTRODUCTION

The problem of a varactor limiter on a $\lambda/4$ stub is treated here on the basis of the equivalent circuit shown in Figure B-1.

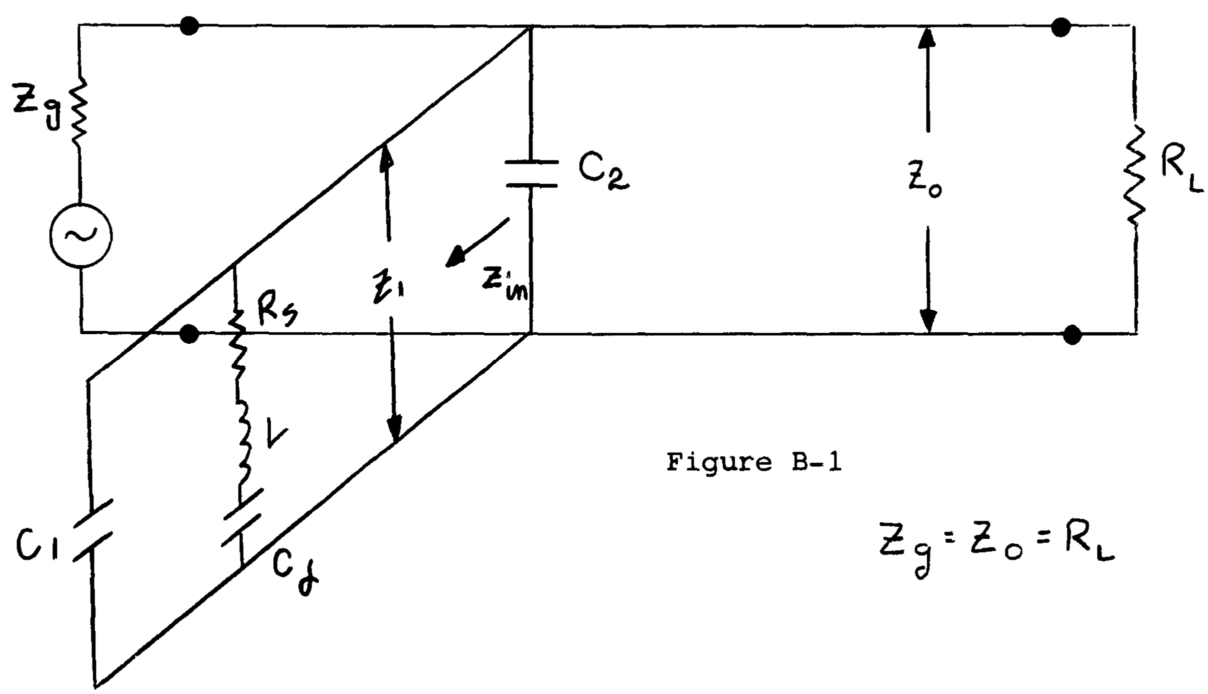


Figure B-1

$Z_g = Z_o = R_L$

C_j = junction capacitance

L = Lead inductance

R_L = Varactor resistance

C_1 = High power tuning adjustment

C_2 = Low power tuning adjustment

Z_T = Impedence of $\lambda/4$ line

Z_{in} = Input impedence looking into $\lambda/4$ line

QUARTER WAVE STUB TERMINATING IMPEDANCE

At high power levels, the junction capacitance C_j is very high. It follows that the capacitive reactance is quite small, and the $\lambda/4$ stub terminating impedance is

$$Z_L \approx \frac{-j \frac{1}{\omega C_1} (R_s + j\ \omega L)}{R_s + j \left(\omega L - \frac{1}{\omega C_1}\right)}$$

Then,

$$Z_{in} = \frac{Z_T^2}{Z_2} = Z_T^2 \frac{R_s + j \left(\omega L - \frac{1}{\omega C_1}\right)}{\frac{L}{C_1} - j \frac{R_s}{\omega C_1}}$$

For best limiting, i.e., small power transmission, Z_{in} should be as small as possible. If C_1 is adjusted so that

$$\omega L = \frac{1}{\omega C_1} ,$$

then

$$(Z_{in})\text{min} = Z_T^2 \frac{R_s}{\frac{L}{C_1} - j \frac{R_s}{\omega C_1}}$$

$$= \frac{Z_T^2 R_s}{-X_L (X_L - jR_s)} .$$

At low power levels, X_{cj} is not negligible so that the $\lambda/4$ stub terminating impedance becomes

$$Z_L' = \frac{\left[R_s + j\left(\omega L - \frac{1}{\omega C_j}\right)\right]\left(-\frac{j}{\omega C_1}\right)}{R_s + j\left(\omega L - \frac{1}{\omega C_j} - \frac{1}{\omega C_1}\right)}$$

But C_1 has been adjusted for the high power case so that

$$\omega L = \frac{1}{\omega C_1}$$

$$\therefore \quad Z_L' = \frac{\left[R_s + j\left(\omega L - \frac{1}{\omega C_j}\right)\right]\left(-\frac{j}{\omega C_1}\right)}{R_s - \frac{j}{\omega C_j}}$$

then

$$Z_{in}' = \frac{Z_T^2}{Z_L'} = Z_T^2 \frac{R_s - \frac{j}{\omega C_j}}{\left[R_s + j\left(\omega L - \frac{1}{\omega C_j}\right)\right]\left(-\frac{j}{\omega C_1}\right)}$$

To obtain minimum loss at low power, Z_{in} should be as high as possible. If the varactor can be chosen so that

$$\frac{1}{\omega C_j} = \omega L ,$$

the circuit will be resonant and C_2 will not be needed. Then

$$Z_{in}' = Z_T^2 \frac{R_s - j\frac{1}{\omega C_j}}{-\frac{j}{\omega C_1}\left[R_s\right]}$$

$$Z_{in}' = Z_T^2 \frac{R_s + j\, X_{cj}}{-j\, R_s X_L}$$

Since the varactor cannot, in gereral, be chosen so precisely for a given circuit, it is necessary to use a tuning capacitor C_2 in order to achieve a resonant condition. The presence of the extra capacitor does not appreciably change the final result.

POWER DISSIPATED IN VARACTOR

The reflection coefficient for the circuit in Figure B-2 is given by

$$\Gamma = \frac{Z_o - \dfrac{R_s R_L}{R_s + R_L}}{Z_o + \dfrac{R_s R_L}{R_s + R_L}}$$

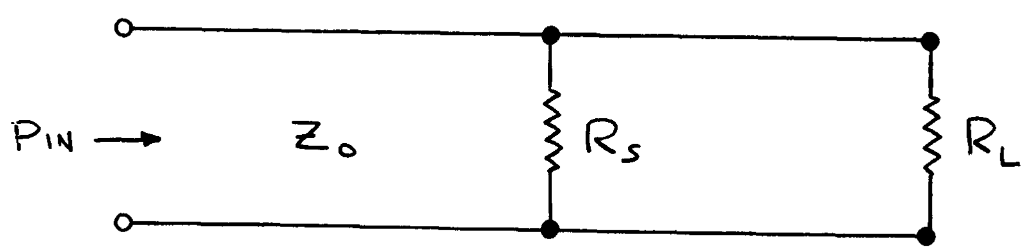


Figure B-2

R_L can be considered to be a matched load and is thus equal to Z_o. Γ then reduces to

$$\Gamma = \frac{Z_o^{\,2}}{Z_o^{\,2} + 2R_s Z_o}$$

Then

$$1 - \Gamma^2 = \frac{4R_s\,(R_s + Z_o)}{(Z_o + 2R_s)^2}$$

The ratios of the power dissipated in the varactor and the load to the input power are given by

$$\frac{P_v}{P_{in}} = (1 - \Gamma^2)\left(\frac{R_L}{R_s + Z_o}\right) = \frac{4R_s Z_o}{(Z_o + 2R_s)^2}$$

$$\frac{P_L}{P_{in}} = (1 - \Gamma^2)\left(\frac{R_s}{R_s + Z_o}\right) = \frac{4R_s^{\,2}}{(Z_o + 2R_s)^2}$$

If the numerator and denominator of the right hand side of each equation are divided by $Z_o^{\,2}$,

$$\frac{P_v}{P_{in}} = \frac{4\,\dfrac{R_s}{Z_o}}{\left(2\,\dfrac{R_s}{Z_o} + 1\right)^2}$$

$$\frac{P_L}{P_{in}} = \frac{4\left(\dfrac{R_s}{Z_o}\right)^2}{\left(2\,\dfrac{R_s}{Z_o} + 1\right)^2}$$

Figure B-3 is a plot of $\frac{P_V}{P_{in}}$ vs $\frac{P_L}{P_{in}}$ (db), showing what fraction of the total power is dissipated in the various levels of limiting.

LIMITING TO LOSS RATIO

It can be shown that

$$\left(\frac{\omega_o}{\omega_c}\right)^2 \approx \frac{\frac{R_s}{Z_o}}{\frac{R'}{Z_o}} = \frac{Z_{in}}{Z_{in}'} \quad \text{for our special case.}$$

Provided that $\omega_c >> \omega_o$.

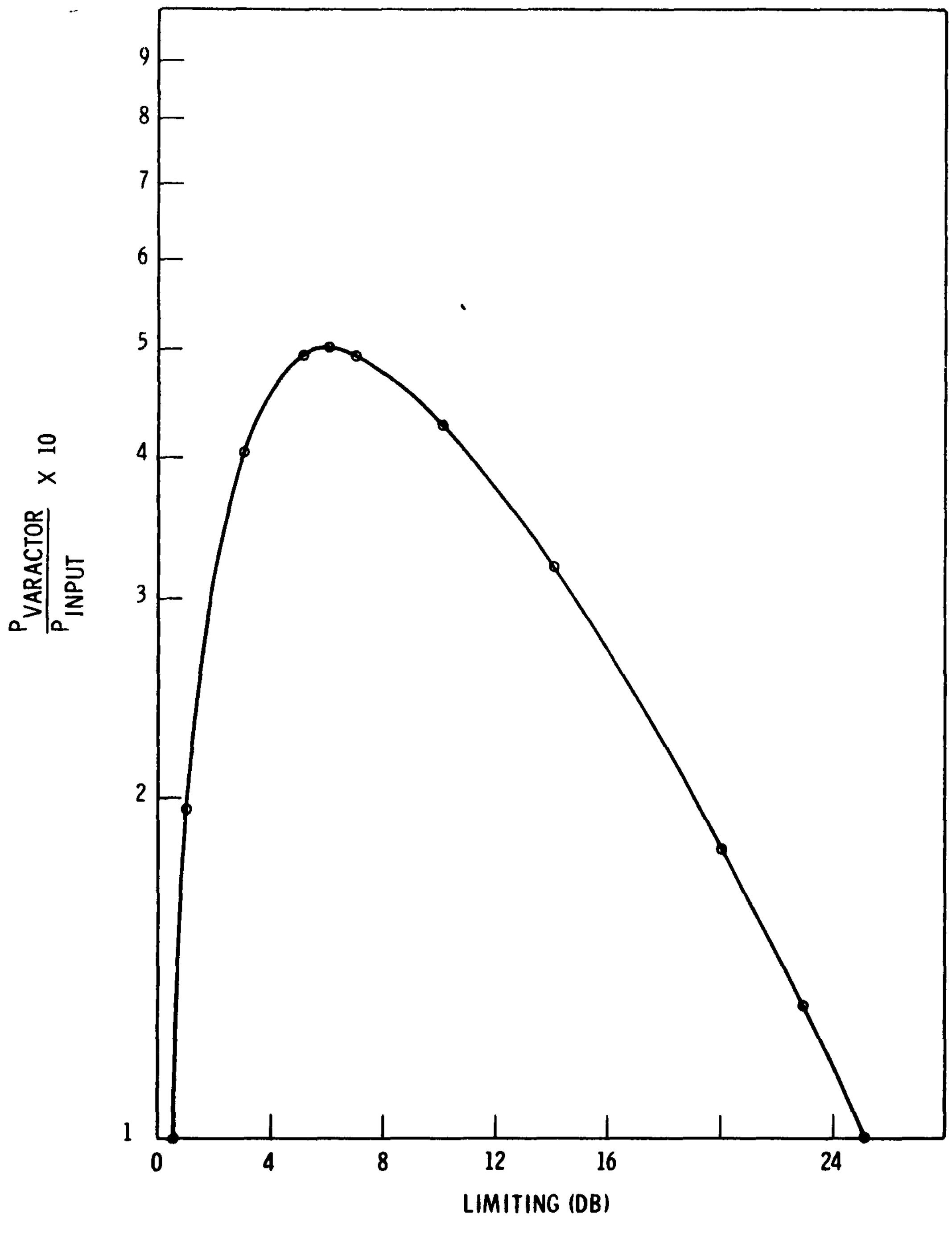
where

ω_o = Operating frequency

ω_c = cutoff frequency of varactor

R' = Low power varactor impedence

Figure B-4 shows $\frac{R_s}{Z_o}$ plotted versus $\frac{P_L}{P_{in}}$ which depicts the effect of the varactor impedance switching ratio on the limiting. $\frac{Z_{in}}{Z_{in}'}$ and thus $\left(\frac{\omega_c}{\omega_o}\right)^2$ is the range covered by $\frac{R_s}{Z_o}$ in going from the high power to the low power case. For



9
8
7
6
5
4
3
2
1
$\frac{P_{VARACTOR}}{P_{INPUT}}$ X 10
0
4
8
12
16
24
LIMITING (DB)

Figure B-3

example if $\left(\frac{\omega_c}{\omega_o}\right)^2 = 1000$ and the low power insertion loss for a given structure is found to be 0.4 db, then the limiting will be 31.2 db.

BANDWIDTH

The bandwidth of the varactor can be found by consideration of the Q of the equivalent circuit in the limiting (high power) case.

$$Q = \frac{|X_L|}{R_s}$$

$$\frac{\omega_o L}{R_s} = \frac{\omega_o}{2\pi BW}$$

$$\therefore \quad BW = \frac{R_s}{2\pi L}$$

THRESHOLD

The limiting threshold of a varactor limiter may be determined by the following analysis:

$$i_v = \sqrt{\frac{P_v}{R_s}} = \text{current in the varactor}$$

$$e_r = \sqrt{R_s P_v} = \text{voltage across the resistance } R_s$$

Since $Q = \frac{X}{R_s} = \frac{1}{\omega_o C R_s}$

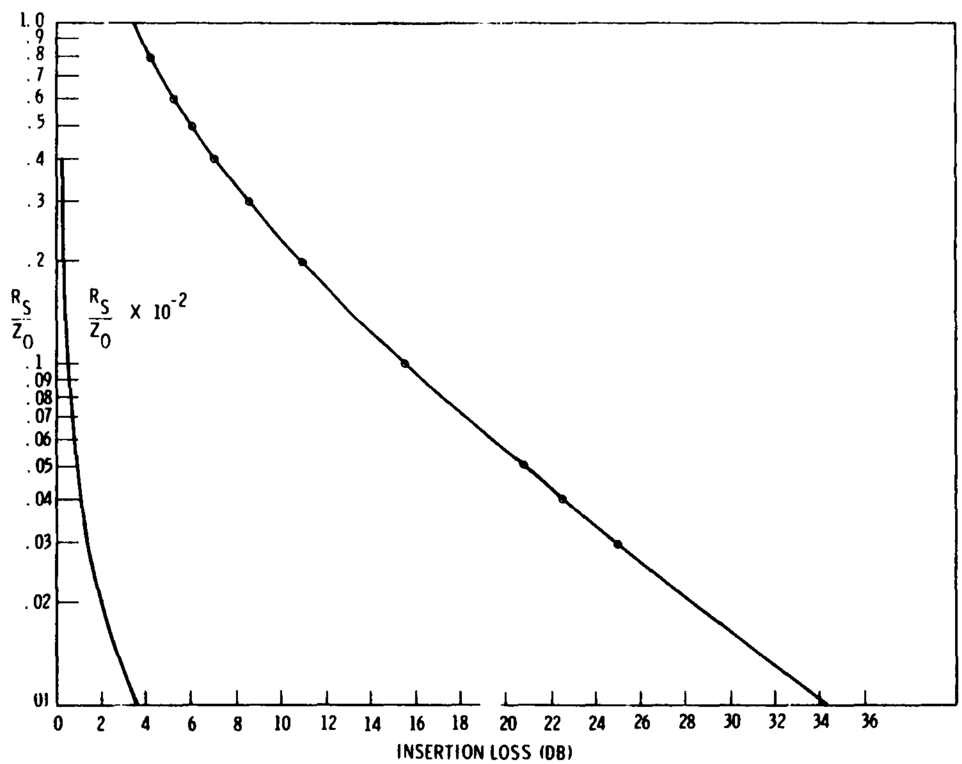

1.0
.9
.8
.7
.6
.5
.4
.3
.2
$\frac{R_S}{Z_0}$
$\frac{R_S}{Z_0}$ X 10^{-2}
.1
.09
.08
.07
.06
.05
.04
.03
.02
.01
0 2 4 6 8 10 12 14 16 18 20 22 24 26 28 30 32 34 36
INSERTION LOSS (DB)

Figure B-4

and $\omega_c = \frac{1}{CR_s}$

$$\frac{\omega_c}{\omega_o} = \frac{X_c}{R_s} = \frac{e_c}{e_r}$$

$$e_c = \frac{\omega_c}{\omega_o} e_r = \frac{\omega_c}{\omega_o} \sqrt{R_s P_v} = \text{voltage across capacitance } C_j$$

The varactor will begin to limit at the input power level for which $e_c = \phi$, where ϕ is the forward conductance voltage of the varactor.

$$\sqrt{R_s P_v} = \frac{\omega_o}{\omega_c} \phi$$

$$P_v = \frac{\phi^2 \left(\frac{\omega_o}{\omega_c}\right)^2}{R_s}$$

The minimum input power for limiting is then

$$\left(P_{in}\right) \min = \frac{1}{4} \frac{R_L}{R_s} P_v$$

$$\left(P_{in}\right) \min = \frac{1}{4} \frac{R_L}{R_s^2} \left(\frac{\omega_o}{\omega_c}\right)^2 \phi^2 .$$

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The goal of this program is a C-band duplexer-limiter package meeting stringent electrical and mechanical requirements. The program has involved principally the investigation of four components: a four port coaxial junction circulator, a coaxial subsidiary resonance limiter, various forms of the multisphere gyromagnetic coupler, and a varactor limiter. Progress on all of these components is reported. A detailed theoretical analysis of the multisphere coupler and experimental results with the varactor limiter are presented in this report for the first time.

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2. Duplexers, C-Band
3. Circulator
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5. Varactor limiter

I. Duplexer-Limiter Package
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